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WIDEBAND, HIGH GAIN, VARIABLE TIME DELAY TECHNIQUES FOR ARRAY ANTENNAS

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ABSTRACT

This report discusses two improved techniques for obtaining a wideband, high gain, nondispersive, variable time delay device to steer array antennas. The first time delay technique uses two wideband helix structures separated by a cylindrical drift tube to obtain a variable delay of 20 or more nanoseconds with a gain of 20 to 30 db. A variable delay is obtained by controlling the beam velocity through the drift region by varying the drift tube potential. Gain variation is minimized by maintaining constant beam velocity through the helices.

The second method is a digital technique in which different time delay elements are switched into the signal path. In this technique, each switch is replaced by a wideband amplifier that determines the system bandwidth (200 to 800 mc).

PUBLICATION REVIEW

This report has been reviewed and is approved.

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WIDEBAND, HIGH GAIN, VARIABLE TIME DELAY TECHNIQUES FOR ARRAY ANTENNAS

INTRODUCTION

The introduction of missiles and satellites into our defense environment has greatly increased the demands placed on modern radar technology. Longer detection ranges of smaller target cross section(s) and higher data rates have forced the designer into the array antenna area. Array systems have the advantage of greater power handling capability and rapid single or multiple beam scanning from stationary structures.

Array antennas have been used for many years in both the communication and radar fields. Until recent years this type of antenna was designed primarily to handle narrowband signals and as such was referred to as a phased array. Their characteristics in this mode have been extensively studied and are well understood. In going to the array system, range resolution has suffered due to the limited bandwidth problems encountered when narrow pulses or pulse compression like linear or quadratic freq-time waveforms are incorporated into the phase steered system.

In array antennas it is desirable to steer the position of the main beam by varying the time delay between elements; however, in narrowband systems it has been shown that steering, by varying the relative phase shift between elements, produces the same result with negligible loss in signal to noise. The relative phase shift method is simple and inexpensive to implement.

As the bandwidth of a phase steered array system is increased, a point is quickly reached where the loss in signal to noise no longer can be neglected. This deterioration in system performance of a phase steered array is caused by two main factors. The first and most important factor is due to the finite propagation or transient time across the array aperture for off-broadside beam positions. For a narrow pulse it is this effect that produces a finite build-up and decay of the signal. For instance, when the pulse length is equal to array transient time, the S/N will be down 3 db. For certain types of pulse compression this finite transit time prevents coherent element-to-element summation off-boresight.

The second factor that can contribute to signal deterioration is the increase in half-power beamwidth (decreased antenna gain) due to an increase in the pulse bandwidth. Assuming wideband elements and phase shifters, this effect is usually of secondary importance and is highly dependent on the feed system used. At most, this effect is about 10 percent of the transient effect degradation. In a series feed array antenna where the phase is inserted between elements (as compared to the parallel feed system), the beam is phased to steer at a single carrier frequency. Power radiated at different frequencies in the spectrum will be steered to different directions. Thus, wider spectrum of transmitted frequencies results in beam broadening. In a parallel feed system (element feed line lengths are the same lengths), the element-to-element phase is also a function of frequency. In this case, the frequency dependence is a function of beam position off-boresight and the type phasing employed. Both of the above problems can be overcome by the proper use of true time delay steering.

Methods have been devised for compensating phased arrays in order that they might handle wideband signals, however, the added complexity makes such techniques questionable. The major component problem in building a wideband array system is the realization of a lossless, wideband, nondispersive, variable time delay device.

Rome Air Development Center sponsored work at present has been limited to the Philco lumped-parameter delay line technique at i-f which utilizes all pass networks similar to the symmetrical lattice phase correction network [18]. Another technique that appears attractive is the cross field technique described by Klüver [1]. Work at Lincoln Laboratory has concentrated on the diode switched digital delay technique at both i-f and r-f [16]. Other proposed digital techniques include the use of new ferrite switches for the higher microwave frequencies.

The purpose of this report is to present two improved techniques for obtaining a wideband variable time delay device to steer array antennas.

VARIABLE TIME DELAY REQUIREMENTS

To time delay steer the beam of an array antenna off-boresight by an angle α, a differential time delay, ΔT, must be introduced into each element or subgroup signal path. It can be shown that the desired differential time delay for the Mth element or subgroup signal path (M 0 is taken as the center element or subgroup with a total count of (2M+1) in the array) is given by the expression:

$$\Delta T + M\Delta T_o \sin\alpha \quad \frac{M\Delta X_o}{c} \sin\alpha , \tag{1}$$

where

c is the speed of light in free space,

ΔT_0 is the propagation time between elements or subgroups, and

ΔX_0 is the element or subgroup spacing.

Notice that the total differential delay required to scan $\pm \alpha$ would be twice ΔT.

If we assume the array is $n\lambda$ long (i.e. $(2M+1)=n\lambda$), then for a scan angle of ± 45 degrees coverage, the maximum delay variation required can be written as

$$\Delta T_{max.} = \frac{0.707 n\lambda}{c} = \frac{0.707 n}{f_0}. \tag{2}$$

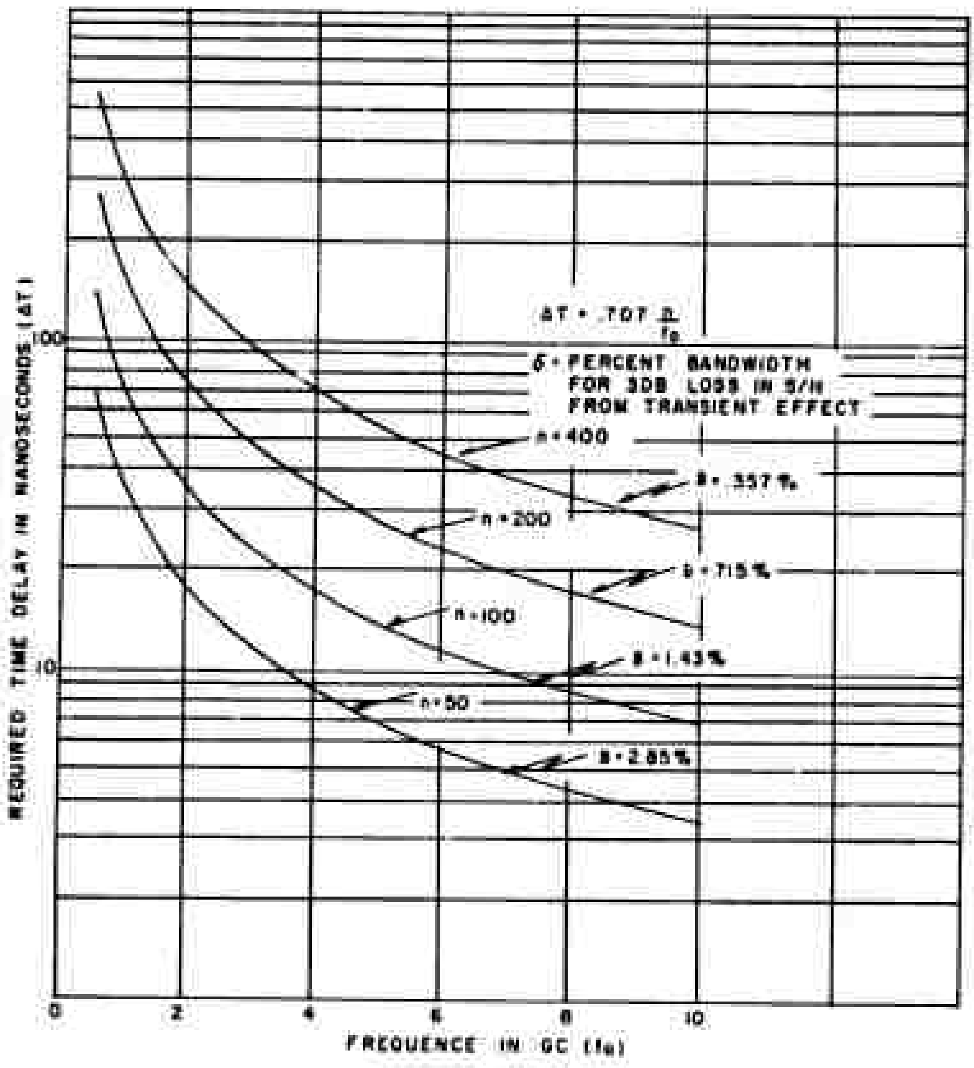
Here the array is scanned about the center element.

This expression gives the maximum delay as a function of both the center frequency of operation and array length n in terms of wavelengths. Figure 1 is a plot of this equation for array lengths of 50, 100, 200, and 400λ as a function of the operating frequency. Also indicated is the percent bandwidth where the signal is down 3 db due to the lack of time delay steering. This is indicated by δ.

As an example, for 100λ array at 4 GC (24.6 ft.) the maximum delay variation required is 17.65 nanoseconds. In the absence of delay steering the signal would be down 3 db for bandwidth greater than 1.43 percent or 57 mcs.

There are two basic ways for obtaining variable time delay. One logical way would be to vary the propagation velocity, v_p, down a transmission or guide structure. If the structure was a conventional waveguide, variable delays could be obtained by operating the guide near cutoff and varying either the physical dimension of the guide or the r-f frequency. However, waveguide structures generally are not satisfactory for wideband signals since they introduce dispersion and loss. Physical movement of components are too slow; however, this method has been used with narrowband signals. The first time delay technique to be discussed below obtains a variable time delay by controlling the signal propagation velocity down a broadband, nondispersive structure. It is also possible to obtain gain at the same time from this device.

Another method for obtaining variable time delay can be realized by switching different time delay elements into the signal path. Such a technique is usually controlled digitally and is referred to as a digital delay line. The second delay technique discussed below realizes a wideband, high gain, variable delay by this digital technique.



ΔT = .707 δ/fo
δ = PERCENT BANDWIDTH
FOR 3DB LOSS IN S/N
FROM TRANSIENT EFFECT
n = 400
n = 200
n = 100
n = 50
δ = .357 %
δ = .715 %
δ = 1.43 %
δ = 2.85 %
100
10
0
2
4
6
8
10
REQUIRED TIME DELAY IN NANOSECONDS (ΔT)
FREQUENCE IN GC (fo)

Figure 1.

VELOCITY CONTROLLED ELECTRON BEAM TIME DELAY

If, instead of trying to vary the propagation velocity of electromagnetic waves down a guiding medium we control the velocity of a modulated electron beam down a drift region, then it appears feasible to build a true variable time delay (delays greater than 20 nanoseconds) device for array antennas. The basic configuration is shown in Figure 2.

An electron beam emitted by the cathode is accelerated toward the first helix by the accelerating potential V_a. The beam, which is confined by magnetic focusing, passes through the input coupling helix at a constant velocity. Upon leaving helix 1, the beam is further accelerated or decelerated by the potential V_b between the helix and drift region. The beam moves through the drift region at its new velocity and is then decelerated or accelerated by $-V_b$ as it passes into the second helix.

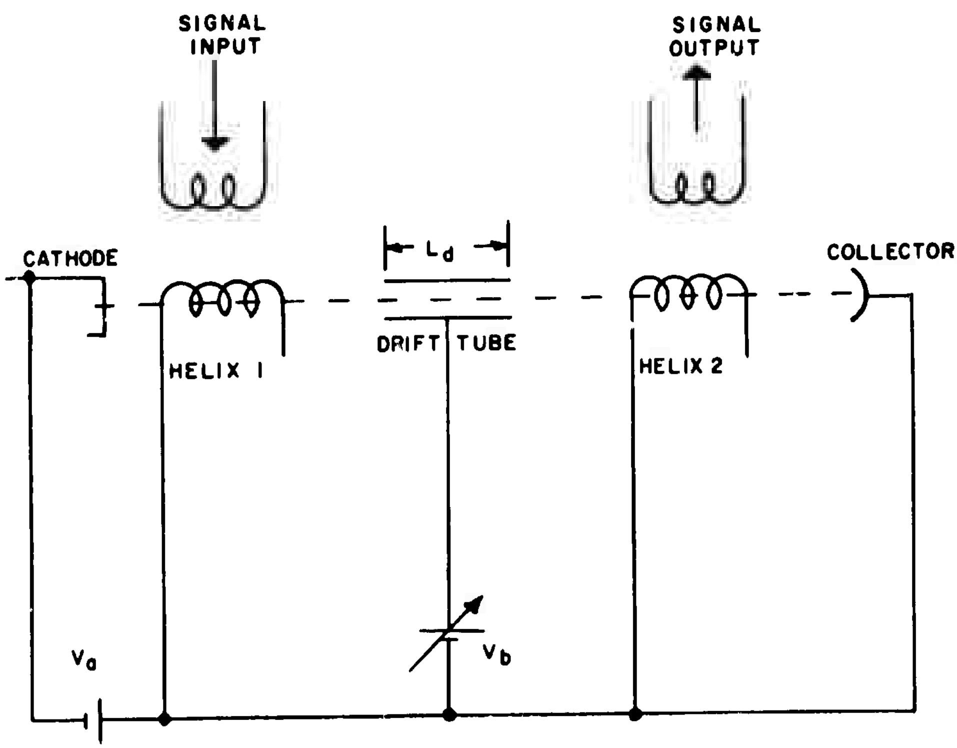


Figure 2. Traveling Wave Variable Delay Tube

Energy coupled onto helix 1 is in turn coupled to the beam. V_a is adjusted so the beam velocity and helix phase velocity are near synchronous. The length of helix 1 is made to correspond to the Kampfner dip length [2]. This is the length for which all of the helix energy is transferred to the beam.

The variable time delay is obtained by varying the beam velocity in the drift region, L_d. The beam velocity, v_b, is a function of the two potentials and can be written as

$$v_b \sim \sqrt{V_a + V_b} = \sqrt{V_d}. \tag{3}$$

Normally, V_a would be held constant and V_b varied.

The time T required for the beam to traverse L_d is

$$T = \frac{L_d}{v_b}. \tag{4}$$

A change in beam velocity through the drift tube will result in a change in delay. That is

$$\Delta T = T_1 - T_2 = \frac{v_{b_2} - v_{b_1}}{(v_{b_1} \; v_{b_2})} L_d . \tag{5}$$

The beam velocity, v_b, is related to the accelerating potential V_d by the expression

$$v_b = \sqrt{\frac{2e V_d}{m}}. \tag{6}$$

Here e/m is the electron charge to mass ratio.

The change in time delay, ΔT, produced by varying the drift tubes potential, from V_{d_1} to V_{d_2} can be written as

$$\Delta T = L_d \sqrt{\frac{m}{2e}} \left[\frac{\sqrt{V_{d_2}} - \sqrt{V_{d_1}}}{\sqrt{V_{d_1} V_{d_2}}} \right],$$

or

$$\Delta T = L_d \; 42.8 \left[V_{d_2}^{-1/2} - V_{d_1}^{-1/2} \right] \text{ nanosecond,} \tag{7}$$

where L_d is in inches and V_d is in volts.

For a 300 volt change (± 150) in V_d centered about 200 volts and a drift length of six inches, a variable time delay of $\Delta T = 23.1$ nanoseconds is obtained. Lower drift velocities will enable large ΔT's to be obtained.

The beam is accelerated or decelerated back to the helix synchronous velocity before entering helix 2. R-f energy will be coupled to the helix. If the helix is long enough, amplification will be obtained.

Since traveling-wave tubes are inherently broadband devices, bandwidths of up to 50 percent should be realizable. The use of TWT structures for phased array steering is not a new concept and has received limited investigation in the past. These techniques were used in narrowband systems where a maximum shift of 360 degrees is all that was required. Larger amounts of phase shifts are redundant and could be eliminated. However, in wideband steering systems time delays greater than that equivalent to 360 degrees of phase shift are necessary. Some of these earlier techniques are briefly discussed below.

The TWT technique employed by Illinois Institute of Technology [3], Buchmiller at Stanford [4], and early Hughes work [5] utilized a conventional TWT and placed a number of equally spaced couplers along the length of the helix. The relative phase shift between adjacent couplers was varied by controlling the d-c helix voltage (beam velocity down the helix). Array phase steering was demonstrated in each case. Although the multiple output (or input) phase shifting technique appears to be quite satisfactory and simple, it has several severe disadvantages. The major problems are: (1) the location of the couplers is very critical; (2) matching each coupler over a wide bandwidth is

very difficult if not impossible; and (3) the gain of the devices vary 6 to 10 db when the array beam is steered off-boresight by 45 degrees. This gain change is a result of varying the beam velocity above and below the synchronous velocity.

A different approach taken by the General Electric Company at C-Band used a split helix TWT amplifier [6]. This low noise amplifier was strictly a two port device in which the helix was separated 0.75 to 2 inches to form a drift region. The drift region was simply an open space between the two halves of the helix. The input helix was maintained at the voltage for optimum noise figure. Phase shift was accomplished by varying the d-c potential between the two helices. This drift region was not a true drift space since the beam experiences a constant accelerating force due to the potential difference between helices. A shift of 360 degrees in phase resulted in a gain variation of 2 to 3 db. This gain variation was primarily due to the variation of beam velocity through the output helix.

A more recent development has been the traveling wave phase shifter developed by Hughes [7] at X-Band. Its basic configuration is identical to the technique depicted in Figure 2. The two helices are maintained at constant potentials and thus no gain variation is experienced due to beam velocity variation within the helix. Again, however, the Hughes device is intended to introduce a very limited variation in phase and thus the drift region is only about 1.5 inch. Figure 3 shows the variation in phase as a function of drift tube voltage for this tube. The gain was measured to be in the range of 35 to 43 db at frequencies of from 8 to 12 GC. The maximum gain variation as a function of drift tube voltage was about ± 1 db with a static drift tube potential of 300 V. It differs from the General Electric Company split helix tube in that a constant potential (constant velocity-zero acceleration) drift region is obtained by surrounding the region between helices with a cylinder whose potential is varied to control the beam velocity through it.

The drift region of the variable time delay tube pictured in Figure 2 is being extended to about seven inches at X-Band. To obtain large delay variations, the lower limits of the drift tube potentials will have to approach the 20 to 30 volt range.

As the variation in delay is increased by either lowering the minimum drift tube potential or by increasing the drift region length, two major problems arise. At low drift tube potentials, beam defocusing becomes important, particularly when the drift region length is increased. This results in excessive beam intercept current on the drift tube cylinder and output helix. The effect can be reduced by increasing the axial magnetic confinement field or by using a beam forming or electrostatic focusing techniques either between the input helix and drift region or within the drift space itself. A simple electrostatic focusing

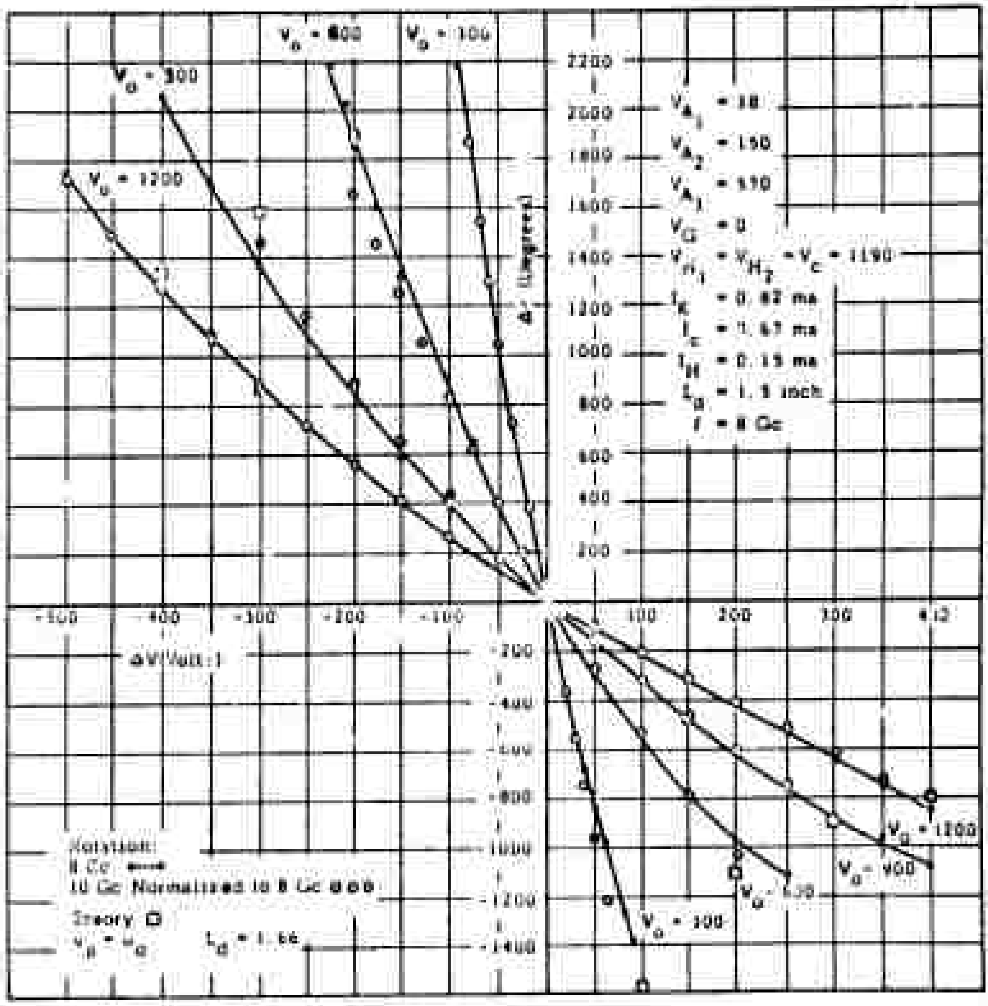


Figure 3. Phase Shift ($\Delta\phi$) as a Function of Incremental Change in Drift-Tube Voltage (ΔV) with the Static Drift-Tube Voltage (V_0) as a Parameter.
solid points - measured phase shift for 8 Gc operation
circles - data points for phase shift at 10 Gc normalized to 8 Gc
squares - computed values

technique within the input helix to drift tube transition region would be to optimize the helix to drift tube diameter ratio. For very long drift spaces the drift cylinder itself could be divided into several lens sections of varying diameter and potentials.

From published experimental results the defocusing problem is not expected to present a major problem. Caulton, Hershenov, and Paschke,[8] in their experiments with Landau Damping, passed electron beams ranging from 50μa to 0.5 ma down a 20 cm drift space with the relatively low drift potential of 50 volts. A magnetic focusing field of 1000 gauss was used.

The second and more serious problem involves the variation of overall gain as a function time delay (beam velocity through the drift space). Ideally, the electron beam, upon entering the drift space, should be density modulated by the input signal. The beam should be bunched with all electrons having the same velocity. Finite velocity modulation of the beam results in two propagating waves (the slow and fast space-charge waves) whose phase velocities are less and greater respectively, than the electron beam velocity. These waves have been termed the Hahn-Ramo waves[9],[10]. A spatially periodic interference pattern results along the beam, with minimums at every half plasma wavelength, if both waves are excited with equal amplitude[11]. Such patterns are clearly indicated from the experimental results of Beam[12], and Caulton, et al[8].

The fundamental plasma wavelength varies with both beam current and drift potential. Therefore, as the delay is varied by altering the drift space potential, the amplitude of the output signal (gain) will vary. Most modulating techniques introduce some velocity spread within the beam. The end objective is to minimize this velocity spread. Density modulation of the beam with a gridded structure introduces minimum velocity spread, however, in the microwave region such a broadband structure is not easily realizable. Beam modulation by a narrow-gap resonator, besides being relatively narrowband, velocity modulates the beam as in a klystron, causing large variations in gain, and therefore, is very undesirable. The broadband helix structure appears to be a reasonable compromise. The helix will density modulate the beam to a degree, however, it will still introduce velocity modulation. Other coupling techniques are being considered but the helix structure appears superior at this time.

Fortunately, but paradoxical, the amplitude of the spatially periodic interference pattern decreases at large distances from the modulating source. Such a phenomena is referred to as Landau Damping[13] and has been analyzed by Dawson[14], and Berghammer[11] and experimentally verified by Caulton, et al[8]. Berghammer has shown that the decay of the velocity modulation is attributable to a spatial Landau damping of the "fast" space-charge wave.

Mihran[15] investigated the propagation of an r-f signal, instituted as density modulations (grid) with some velocity modulation present. He found that the amplitude of the spatial interference pattern could be controlled by changing the magnetic focusing field when the beam was viewed 15 to 20 cm from the modulating point.

From the preceding considerations a delay variation in excess of 20 nano-seconds with maximum fluctuations in gain of several db are considered realizable by this technique of an extended drift space region. Once the bunched beam enters the second or output helix useful gain of 20 to 30 db can be obtained.

WIDEBAND DIGITALLY CONTROLLED DELAY LINE FOR PHASED ARRAY STEERING

Digitally controlled delay lines or phase shifters have been used for the past several years in steering phased array antennas. Lincoln Laboratory has conducted an extensive program in the development of such devices. Much of Lincoln's work can be found in their technical reports[16].

Most digital delay line or phase shift devices use diodes to switch the signal through or around a length of transmission line cut to give the desired incremental delay or phase shift. A commonly used technique is shown in Figure 4. The input signal can be delayed through path 1 simply by closing S_1, S_2, and opening S_3, S_4. Path 2 is selected by closing S_3, S_4, and opening S_1, S_2. However, high speed switching requires the elimination of mechanical switching devices. Switching devices in the past have utilized simple diodes with small forward impedance and a high back impedance. Replacing S_1, S_2, S_3, S_4, with diodes D_1, D_2, D_3, D_4, and controlling the magnitude and direction of current flow through the diodes one is able to direct the input signal through either path 1 or path 2.

These diodes, however, have a finite series resistance when forward biased and therefore produce loss. Additional loss is usually introduced so as to reduce the fluctuation in amplitude as the signal path is changed. The matching of these diodes and the proper termination of the incremental delay units is accomplished at the expense of bandwidth. Series diode resistance plus diode reactance also results in poor matching and SWR between the incremental delay units.

A typical 6 bit, 28 mcps Digital phase shifter or delay line is described in Lincoln Laboratory Technical Report 228[16]. It was designed to produce a total of 360 degrees phase shift at 28 mcs by switching either through or around six lengths of coax by the use of diode switches. Total insertion loss was 11 db. The set up time ran about 3 μsec with a maximum amplitude error of 0.25 db and phase error of 2 degrees. Input power capacity was limited to one watt. Each diode switch required about 40 ma to reduce the diode series resistance to a usable value.

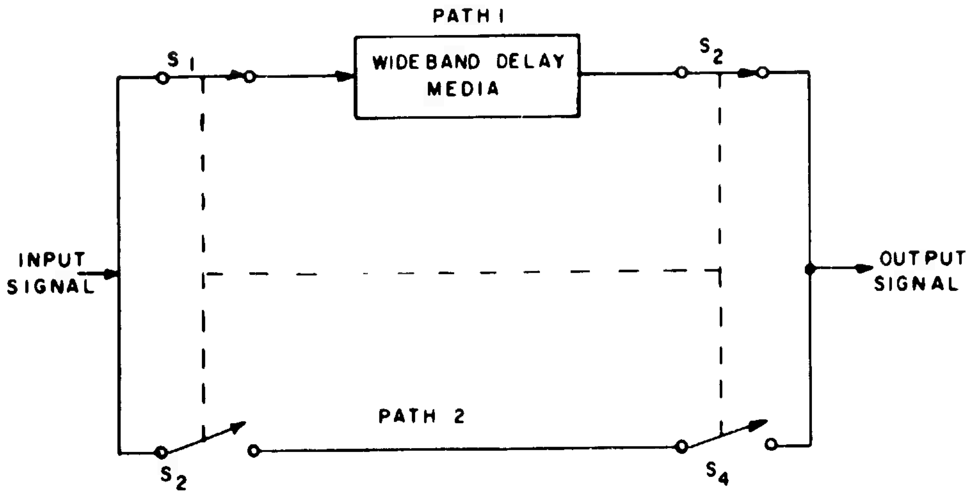


Figure 4. Single Bit Digital Delay Line

The Wideband Digital Delay Line to be described uses the basic technique indicated by Figure 4. However, in this case each mechanical switch is replaced by a wideband amplifier. It is the bandwidth of these amplifiers that determines the system bandwidth (200 to 800 mc). These amplifiers can be gated on or off. By using wideband amplifiers in place of the diode switches three advantages are gained: (1) each delay line (coax) is properly terminated into its characteristic impedance over the frequency band of interest; this prevents loss due to excessive SWR; (2) each delay line increment is isolated from the next due to the presence of amplifiers. (3) with the inclusion of amplifiers the total input-output gain can be made equal to or greater than unity, gains as high as 100 db for a 10 bit device are achievable with bandwidths of 200 mc; however, 100 db is more than normally required with such bandwidths.

The wideband digital delay line is inherently a low frequency device. By low frequency we mean i-f frequencies up to 600 or 800 mc. If we wish to process a signal with a 200 or 300 mc bandwidth the i-f center frequency would more than likely be located in the range of 200 to 300 mc.

High Gain Wideband Digital Delay

The basic building block for the high gain wideband delay device is a stable wideband amplifier that can be gated on or off. One such amplifier will replace each diode switch of Figure 4. The wideband amplifiers to be described used transistors operated in the common base mode, however, vacuum tubes could be used if necessary.

The basic common base amplifier configuration is shown in Figure 5. The input signal power is:

$$P_{in} = Z_{in}(\Delta i)^2 , \tag{8}$$

where Z_{in} is the input impedance which is approximately equal to r_e ($r_e \sim 25$ to 100 ohms). In the grounded base configuration the emitter signal current is approximately equal to the collector signal current. Therefore, the amplifiers output signal power is:

$$P_{out} = R_2 (\Delta i)^2 . \tag{9}$$

The amplifiers power gain then becomes

$$G_p = \frac{R_2}{Z_{in}} \tag{10}$$

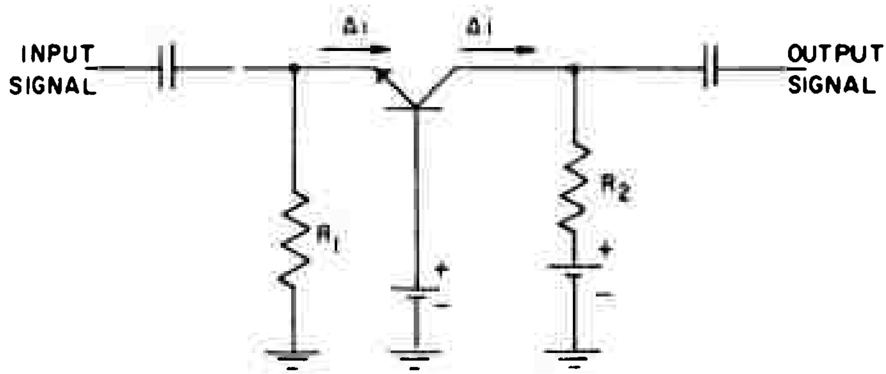


Figure 5. Ground Base Amplifier

The amplifier gain is thus fixed by the ratio of the input impedance to output load. This relationship is valid as long as the cutoff frequency of the transistor is not exceeded. The figure of merit for amplifiers is normally given as the gain X bandwidth produce, f_t, in mcs. Thus, the upper bandwidth limitation, f_c, on the grounded base amplifier can be written as

$$f_c = \frac{f_t}{G_p} . \tag{11}$$

The amplifier can easily be gated off by either grounding the base directly or by applying a small negative bias to the base. When reverse biased, the opened switch produces at least 25 db isolation in the band of interest.

In this particular application it is desirable to have the amplifiers input and output impedance matched to a transmission line of characteristic impedance Z_o. When resistive coupling is used as shown in Figure 5, such that $Z_o = Z_{in} = R_2$, unity gain is obtained from Equation 10. Since longer lengths of coax cable used for delay lines produce attenuation, it is desirable that this amplifier produce gain instead of having to add additional amplifiers at the delay units output. Gain can be obtained by increasing the load impedance seen by the collector. Such an increase in load impedance can be obtained by using broadband transformers like those described by C. L. Ruthroff(17) to transform Z_o. The particular transformer useful in this case is the 4:1 impedance transformer shown in Figure 6, which is wound to form a transmission line. Ruthroff obtained over 700 mcs bandwidths at the 3 db points with the lower cutoff at 200 kcs and the upper cutoff at 715 mcs. If this transformer was combined with the grounded base amplifier of Figure 5, a power gain of 4 (6 db) would be possible when the amplifiers input and output are matched to Z_o. Figure 7 is a schematic of the wideband amplifier.

The transistor shown, 2N709 (Fairchild), has a typical f_t of 800 mcs. The input impedance Z_{in} was measured at about 31 ohms. With the 4:1 impedance transformer coupled to 52 ohms, a bandwidth of over 200 mcs has been observed. This was verified with several 2N709 transistors, which indicates the devices tested had f_t's better than the typical. When the input signal was taken directly from the sweep generator output the response to 200 mcs was within 1 db of being flat. A 10 foot length of 52 ohms coax was then connected between the generator and amplifier input. No attempt was made to match the cable. The amplifier output showed ripples of about 2 db amplitude due to the nonideal termination. The 3 db dropoff was above 250 mc.

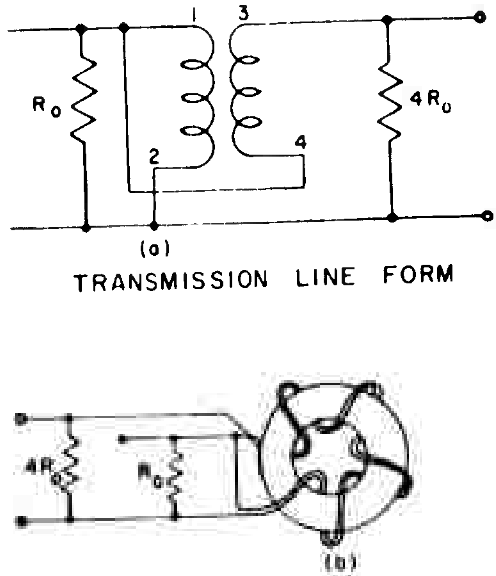


Figure 6. Wiring Diagram for 4:1 Impedance Transformer

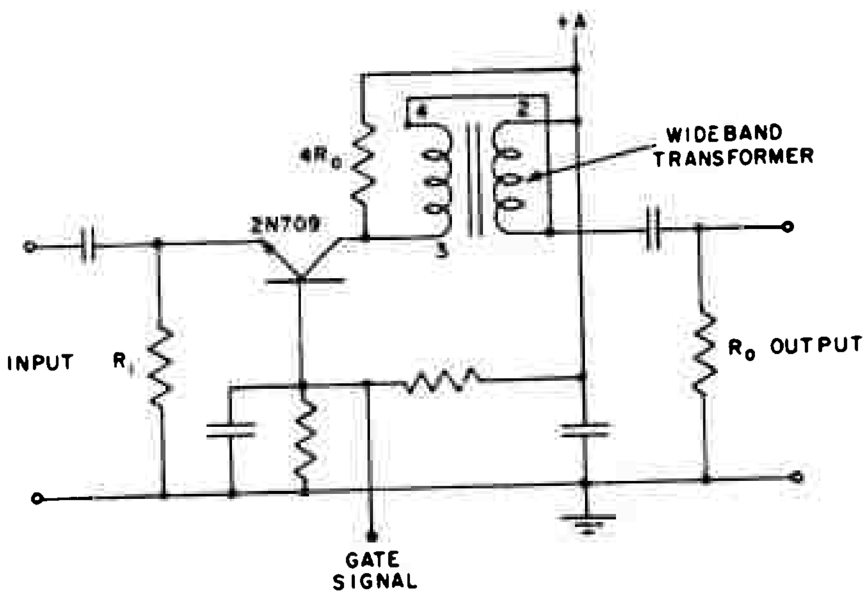


Figure 7. Wideband Amplifier

Since loss will vary from bit to bit due to the different lengths of delay line the amplification must be controlled without disturbing the line match. Gain variation can be accomplished by several techniques. Since the input impedance to the transistor is smaller than Z_0, a small resistance pad added to the input would control the gain as well as the termination of the line. This technique should be avoided, if possible, by either using a transistor with higher grounded base input impedance or by using a lower impedance transmission line. Increasing the input to the transistor by the addition of a series resistor tends to reduce the upper cutoff frequency.

A second and more profitable technique for lowering the amplifiers gain is to load the transformer and thus reduce the amplifiers gain. If either side of the transformer is loaded the other must be loaded accordingly to maintain the 4:1 impedance. For maximum bandwidth it is important that this 4:1 match be maintained. However, a small mismatch will produce a slight rising or falling with frequency that can be used to cancel the attenuation of the delay line as a function of frequency. If each stage gain were reduced to 5 db, a 10 bit delay line would produce 100 db (two amplifiers per bit) of gain over the band required. In a wideband system the overall gain would depend on the bandwidth and front end noise figure. If only 50 db of gain were required, one stage could be operated at unity gain without the transformer. If less bandwidth is desired, a filter could be placed at the input and output of the overall delay unit. The bandwidth of each bit could be extended at the expense of gain. As new transistors become available the bandwidth will be extended for a fixed gain. Devices presently available include 2N2857-f_t~1200; 2N2808A-f_t~1500. These devices are fairly expensive, however. The Philco 2N2966 claims a typical gain of 10.5 db at 800 mc, and is relatively inexpensive.

It should be noted that this digital delay and amplifier device does not require intricate tuning. The transformer is easily wound and the 2N709 transistor is not expensive. If a bandwidth of 150 mcs is desired, a 2N708 could be used at a cost of about $2.00 each which is quite reasonable for the job it performs.

Four such wideband amplifiers connected, as shown in Figure 8, would form one bit of the digital delay device. Notice that a common emitter resistor is used for both inputs. Since both Q_1 and Q_3 are not on at the same time, a single emitter resistor provides better matching of the preceding stage as well as fewer components, shorter stray lead length and capacitance which could reduce bandwidth and increase S.W.R. Also, it should be noted that the output transistors, Q_2 and Q_4, have their collectors connected. Since only one of these stages is on at a time this also provides better matching to the following delay bit stage. As before, this reduces component and stray lead inductors and capacitance. The gain of each path (that is Q_1 or Q_3) can be adjusted by changing R_1 and R_2. The delay bit stage gain can be controlled by introducing resistance, R_4, in series

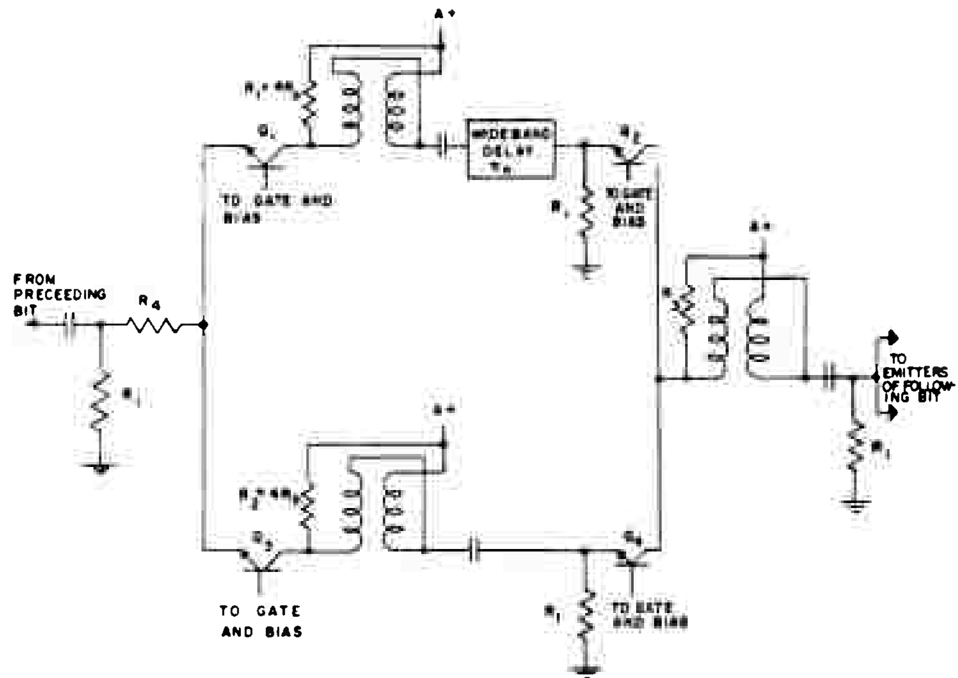


Figure 8. Schematic Diagram of Single Bit of Wideband Digital Delay Device

with the input or by changing R_3. It can be seen that if the digital delay line device produced sufficient gain then an additional wideband i-f amplifier is no longer required. This would reduce the cost per element.

The angular beam coverage generally required is ± 45 degrees or a total of 90 degrees. If the array is 100λ then the 3 db beamwidth would be about 0.5 degrees. If the beam were moved in 0.1 degree increments the resultant scanning would closely resemble a continuous scanning system. This would require 900 beam positions. If each digital delay line had 10 bits this would give 1024 discreet positions. Nine bits would give 512 positions or about 0.2 degree incremental beam steps.

Wideband Delay Elements

Previously, the wideband delay line for insertion between amplifiers Q_1 and Q_2 had not been considered. Any delay device considered should not be too lossy; however, this is of secondary importance with the addition of amplifiers. Of primary importance is the dispersion of these devices. They should be operated well in the nondispersive region. The most obvious method for realizing a wideband delay line would be the use of a fixed length of coax cable. Different lengths of cable would produce different delay times.

Over a wide frequency range, the attenuation of coaxial cable is a function of frequency. For instance RG-8/U which has a characteristic impedance of 52 ohms and 29.5 pfd/ft has an attenuation per 100 feet of 0.16 db at 1 mc; 0.55 db at 10 mc; 2 db at 100 mc; and 8.6 db at 1 GC. A second consideration would include the size of the cable since a 50 foot coil could take up quite a bit of room. There are some manufacturers that produce small low capacitance cable.

A coaxial cable with a high dielectric constant is desirable since this will reduce the propagation velocity and thus result in shorter lengths of cable. RG-8/U has a relative velocity of propagation of 65.9 percent. By mismatching the wideband transformer, it can be made to give a slight increase in gain with frequency. This effect could be used to compensate to a limited degree the attenuation characteristics of the cable. Generally, helical high-impedance delay cables (center conductor wound like a helix) are dispersive above a critical frequency. This upper critical frequency is seldom above 10 mc which would eliminate it from consideration. However, variations of this technique could no doubt be devised to extend this frequency to several hundred megacycles.

The Philco Corporation under Rome Air Development Center sponsorship has been developing an electronically variable time delay technique using an all-pass network similar to the symmetrical lattice phase correction network. Although Philco has not published a description of their device a similar delay technique has been described by R. W. Colfee [18]. The lumped-parameter delay line technique appears attractive as a delay device. A typical network is capable of producing delays from 8 to 15 nanoseconds or a variation of 7 nanoseconds with a bandwidth of 30 to 40 mc. The variation in delay is produced by replacing the fixed capacitors with matched varactors. To extend the bandwidth to several hundred megacycles requires high quality diodes that become quite expensive. As the frequency is increased the delay is decreased approximately and proportionally. If the expensive varactors were replaced by small adjustable capacitors, the total delay through the device could be used to obtain a desired incremental delay. With adjustable capacitors each increment of delay could easily be adjusted to the proper value. The coaxial cable technique could be combined with the adjustable lumped constant network. This would reduce the complexity of final adjustment by providing a simple method of line stretching.

CONCLUSIONS

Two variable time delay devices have been discussed. Both techniques, although still in the construction stage, appear capable of producing the required delay variation with a large gain and bandwidth.

The TWT technique has the advantage of (1) operating at the radiating frequency; (2) octave instantaneous bandwidth capabilities; (3) producing gain (at moderate noise figures) and delay within the same envelope; (4) occupying a relatively small space behind each element; and (5) large dynamic range. If the beam current were increased (incompatible with long delays) then such a tube could be used in the transmit array. Such a device would probably be used to time steer subgroups of elements instead of one per element. With appropriate high power amplifiers in each subgroup, it is desirable that the TWT delay amplifiers provide at least several kilowatts of peak output power.

Disadvantages of the TWT technique include (1) high power and multivoltage sources; (2) requirement for excellent voltage stability, particularly for the drift tube potential; and (3) heat dissipation problem. To maintain good phase stability, especially if the tube is operated at higher power levels, a closed loop stabilization system should be considered.

The digital delay technique is capable of much better phase stability since fixed delays are required in each bit. About the only parameter requiring control to maintain phase stability is temperature. In addition to the phase stability, this delay technique has the advantage of (1) very low power requirements; (2) minimum heat dissipation; and (3) high gain and wide bandwidths.

Disadvantages include (1) limited power handling capacity (this is primarily a receiving technique but could be used in the transmitter with appropriate power gain following it); (2) operation at i-f; if radiated frequency is above 1 GC, additional mixers and local oscillator phase shifters are required, and (3) larger in size than the TWT, however, power requirements are considerably reduced.

With either or both of these variable delay techniques, wideband time steered array antenna systems capable of handling any type of signal waveform (short pulse, chirp, etc.) would be possible.

REFERENCES

1. Klüver, J. W., "An Electronically Variable Delay Line," Proceedings IRE, Vol. 50, p 2487; December 1962

2. Hutter, R. G. E., "Beam and Wave Electronics in Microwave Tubes," Van Nostrand Company, Inc., p 332, 1960

3. Cohn, G. I., "Theoretical Study and Investigation of Techniques for an Electronic Rapid-Scan Antenna," Phase I, Ia, II, Electronics Research Laboratory Illinois Institute of Technology, Contract No. DA-36-039 SC-70114, April 1, 1959

4. Buchmiller, L. D., "A Multiple Input Traveling Wave Tube Phase Shifter," Stanford University, Tech. No. 558-1, Contract No. DA-36-039-SC-78296, October 10, 1960

5. "High Speed Electronic Antenna Scanning in the Microwave Region," Hughes Aircraft Co.; Tech. Report RADC-TR-61-197, AF30(602)-2156, April 1961

6. "Low-Noise C-Band Traveling-Wave Amplifier," General Electric Company; RADC-TDR-62-445, AF30(602)-2436, September 11, 1962

7. "Traveling-Wave Phase Shifter," Hughes Aircraft Co., RADC-TDR-63-498, AF30(602)-3005, December 1963 (Confidential)

8. Caulton, M., Hershenov, B., and Paschke, F., "Experimental Evidence of Landau Damping in Electron Beams," J. Appl. Phys., Vol. 33, pp 800-803, March 1962

9. Hahn, W. C., "Small Signal Theory of Velocity-Modulated Electron Beams," General Electric Rev., Vol. 42, pp 258-270; 1939

10. Ramo, S., "Space Charge and Field Waves in an Electron Beam," Phys. Rev., Vol. 56, pp 278-283; 1939

11. Berghammer, J., "Landau Damping of Space-Charge Waves," J. Appl. Phys., Vol. 33, pp 1499-1504; April 1962

12. Beam, W. R., "On the Possibility of Amplification in Space-Charge-Potential-Depressed Electron Streams," Proc. IRE, Vol. 43, pp 454-462; April 1955

13. Landau, L., J. Phys, U.S.S.R., Vol. 10, p 25; 1946

14. Dawson, J., "On Landau Damping," Phys. of Fluids, Vol. 4, pp 869-874; July 1961

15. Mihran, T. G., "R. F. Current Behavior in Electron Beams with dc Velocity Spread," J. Appl. Phys., Vol. 33, pp 1582-1590; April 1962

16. "Phased Array Radar Studies," Lincoln Laboratory, Tech. Reports 228 (August 1960), AD-249 470; 236 (November 1961), AD-271 724; 299 (February 1963)

17. Ruthroff, C. L., "Some Broad-band Transformers," Proc. IRE, Vol. 47 pp 1337-1342; August 1959

18. Calfee, R. W., "An Active Network Equivalent to the Constant-Resistance Lattice with Delay Circuit Applications," IEEE PTGCT, Vol. p 532; December 1963